REMARKS

Claims 1-6 and 9-21 remain pending in the present application. Applicant thanks the Examiner for the allowance of claim 6. Consideration of pending claims 1-5 and 9-21 and allowance of these claims is respectfully requested in view of the following comments.

Claim Rejections pursuant to 35 U.S.C. §112, first and second paragraphs

Claims 9-10 were rejected pursuant to 35 U.S.C. §112, first and second paragraphs. The Examiner has deemed the connection between the output of the demodulation filter and the feedback control loop as critical or essential to the practice of the invention. Applicant respectfully refers the Examiner to paragraph 11 on page 3 in which a power amplifier system is described that includes a feedback control loop having a feedback demodulation filter. Those skilled in the art would recognize that a circuit with a feedback control loop that includes the feedback demodulation filter as disclosed by claim 9 and described by paragraph 11 would allow operation of the power amplifier system, and is therefore enabled by the specification. Applicant has added a new paragraph in the detailed description that is similar to paragraph 11 on page 3. Based on the power amplifier system described in paragraph 11, Applicant has added existing element numbers in the new paragraph and a corresponding new proposed drawing as FIGURE 6. Applicant has submitted herewith a proposed new drawing to the Official Draftsperson.

Applicant respectfully request approval of the addition of new FIGURE 6 and the new paragraph to the application. A marked up version of paragraph 11 is included as an attached VERSION WITH MARKINGS TO SHOW CHANGES MADE to indicate the differences between the new paragraph and paragraph 11. No new matter has been added by the addition of the new paragraph and FIGURE 6. Since the specification and drawings now even more clearly enable the power amplifier disclosed in claims 9 and 10, Applicant respectfully requests the

removal of the rejection of claims 9 and 10 pursuant to 35 U.S.C. §112, first and second paragraphs.

Claim Rejections pursuant to 35 U.S.C. §112, second paragraph

Claims 3 and 19 were rejected pursuant to 35 U.S.C. §112, second paragraph as being indefinite. The Examiner has indicated that claims 3 and 19 are indefinite since the low-pass filter circuit is not a Sallen & Key filter. Applicant has amended claim 3 to indicate that low-pass filter circuit includes a Sallen & Key filter. Claim 19, however, already discloses that the low-pass filter circuit includes a Sallen & Key filter. The term "includes" does not indicate that the low-pass filter is a Sallen & Key filter, only that the low-pass filter comprises a Sallen & Key filter. The term "includes" is synonymous with the term "comprises." Hewlett-Packard Co. v. Repeat-O-Type Stencil Manufacturing Corp., Inc., 123 F.3d 1445, 1451 (Fed. Cir. 1997). Accordingly, Applicant respectfully requests the removal of the rejection of claims 3 and 19 pursuant to 35 U.S.C. §112, second paragraph.

Claim Rejections pursuant to 35 U.S.C. §102(b)

Claims 1, 2, 4, 5, 13-18, 20 and 21 were rejected pursuant to 35 U.S.C. §102(b) as being anticipated by Cavigelli (U.S. Patent No. 5,635,871 hereinafter referred to as "Cavigelli"). Applicant respectfully disagrees for at least the following reasons.

Claim 1 discloses an active low-pass filter system that includes a low-pass filter circuit and an isolated-integrator band-reject filter. The low-pass filter circuit includes a resistive forward signal flow branch. The isolated-integrator band-reject filter is imbedded within the low pass filter circuit and forms part of the resistive forward signal flow branch. Claim 13 similarly discloses an active low-pass filter system that includes a low-pass filter circuit having an input terminal and an output terminal. Incorporated into the low-pass filter circuit between the input and output terminal is an isolated-integrator band-reject filter.

Cavigelli teaches, in Fig. 1 and 12a a low pass filter with three amplifying stages 7, 13 and 19 cascaded together. In Fig. 12a, in addition to the three cascaded amplifying stages 7, 13 and 19 a cascaded notch filter 202 is also taught. The Examiner has postulated that the cascaded notch filter 202 is equivalent to the isolated integrator band-reject filter disclosed in claims 1 and 13. As defined in Exhibit A, notch filters are a broad class of filters which are frequency rejection circuits, such as a band-suppression filter for producing a notch. Stan Gibilisco, The Illustrated Dictionary of Electronics, p. 474-475 (7th ed., McGraw-Hill, 1997). (see also in Exhibit A, Lawrence P. Huelsman, Chap. XIII Filter Characteristics, Section 65 General Characteristics of Filters, p. 2156 - 2159 (Wai-Kai Chen, The Circuits and Filters Handbook, CRC press, 1995)). Exhibit B includes excerpts from a treatise entitled Electronic Filter Design Handbook, Arthur B. Williams, p. 6-19 through 6-39 (McGraw-Hill 1981), and Active Filter Design Handbook, George S. Moschytz and Peter Horn p. (John Wiley, 1981) to illustrate some of the wide variety of different filters within the class described as notch filters.

In Cavigelli, the notch filter 202 is described as a filter with a notch frequency substantially above the operating frequency that is capable of reducing high frequency noise. (Col. 8 lines 63-67 and Col. 9 lines 1-6) Cavigelli, however, does not teach suggest or disclose that the notch filter 202 is an isolated integrator band-reject filter as disclosed in claims 1 and 13. In fact, Cavigelli fails to disclose any details about the notch filter 202. In Fig. 12a, the notch filter 202 is illustrated as an empty box which clearly does not teach, suggest or disclose anything with regard to the circuit configuration of the notch filter 202. Further, neither the notch filter 202 or any of the amplifying stages 7, 13 and 19 of Cavagelli disclose a tuning resistor as disclosed in claim 2; a resistive forward signal flow branch as disclosed in claim 14; a first resistor and a second resistor with the isolated integrator band-reject filter connected therebetween as disclosed by claim 16; a resistive value of zero as disclosed by claim 17; or at least three capacitors and at least two resistors of equal value as disclosed by claim 18. In fact,

VERSION WITH MARKINGS TO SHOW CHANGES MADE

Please add the following new paragraph immediately following paragraph 46 in the DESCRIPTION OF THE PRESENT INVENTION:

-- As illustrated in Figure 6, t[T]he present invention also provides a power amplifier system 40 for driving a load 45. The power amplifier system 40 includes an input terminal for receiving a signal from a signal generator, an output terminal connected to the load 45, a pulse width modulation circuit 42 creating ripple spectra, an error amplifier and modulator 43 connected to an input of the pulse width modulation circuit 42, a demodulation filter 47 connected to an output of the pulse width modulation circuit 42, and a feedback control loop coupled to the pulse width modulation circuit 42 and including an active low-pass filter having a feedback demodulation filter 44 and an isolated integrator frequency-rejecting network. In one embodiment, the isolated-integrator frequency-rejecting network is an isolated-integrator band-reject filter. --

Please amend Claim 3 as follows:

3. (Amended) The system of Claim 1, wherein the low-pass filter circuit [is]includes a Sallen & Key filter.

none of the cited prior art references teach, suggest or disclose the isolated integrator band-reject filter disclosed in claims 1 and 13, or any of the elements disclosed by claims 2, 14, 16, 17 or 18.

Accordingly, for at least the foregoing reasons, Applicant respectfully requests the Examiner to remove the rejection pursuant to 35 U.S.C. §102(b) of independent claims 1 and 13 and dependent claims 2, 14, 17 and 18. Alternatively, since dependent claims 2-5 and 14-22 depend from respective independent claims 1 and 13, removal of the 35 U.S.C. §102(b) rejection of these dependent claims is respectfully requested.

Applicant believes that claims 1-6 and 9-21 are allowable in their present form and that this application is in condition for allowance. Accordingly, it is respectfully requested that the Examiner so find and issue a Notice of Allowance in due course. Should the Examiner deem a telephone conference to be beneficial in expediting allowance of this application, the Examiner is invited to call the undersigned attorney at the telephone number listed below. No fees are believed to be due at this time, however, should any fees be deemed required, please charge such fees therefor to Deposit Account No. 23-1925.

Respectfully submitted,

Sand N. Hill

Sanders N. Hillis
Attorney Reg. No. 45,712

Attachments: VERSION WITH MARKINGS TO SHOW CHANGES MADE pg. 11
Exhibits A (12 pages)
Exhibit B (28 pages)

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PG 1 OF 4

The Illustrated Dictionary of Electronics

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Editor-in-Chief

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PG 2 OF 4

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PAGE 3 OF 4

normal-distribution curve See BELL-SHAPED CURVE.

normal electrode A standard electrode used in electrode-potential measurements.

normal fault An unintended path between the hot terminal of a load and ground.

normal fault plus grounded neutral fault A combination of NORMAL FAULT and GROUNDED NEUTRAL FAULT.

normal glow discharge In a glow-discharge tube, the discharge region between the Townsend discharge and the abnormal glow in which current increases sharply, but a constant voltage drop is maintained across the tube.

normal impedance A transducer's input impedance when the load impedance is zero.

normal induction curve A saturation curve for a magnetic material. Also see BOX-SHAPED LOOP and SATURABLE REACTOR.

normalize In computer programming, to use floating-point numbers to modify the fixed-point part of a number so that it is within a desired range.

normalized admittance The quantity $1/Z_n$, where Z_n is NORMALIZED IMPEDANCE.

normalized frequency The unitless number represented by the ratio f/f_r, where f_r is a reference frequency and f is a frequency of interest. Response plots are sometimes conveniently drawn on the basis of normalized frequency, the reference (or resonant) frequency being indicated as **1**, twice the reference frequency as **2**, etc.

normalized impedance A value of impedance divided by the characteristic impedance of a waveguide.

normally closed Abbreviation, NC. Pertaining to a switch or relay whose contacts are closed when the device is at rest. Compare NORMALLY OPEN.

normally open Abbreviation, NO. Pertaining to a switch or relay whose contacts are open when the device is at rest. Compare NORMALLY CLOSED.

normal mode Pertaining to a device or system operated in its usual or most common manner.

normal mode A state of acoustic resonance in an enclosure, such as a speaker cabinet or a room.

normal-mode rejection Abbreviation, NMR. In a digital direct-current voltmeter, the level of noise on the applied voltage that will be rejected by the instrument. Compare COMMON-MODE REJECTION.

normal position In a switch or relay, the state of the contacts when the device is at rest.

normal solution A solution, such as an electrolyte, in which the amount of dissolved material is chemically equivalent to 1 gram-atomic weight of hydrogen per liter of the solution. Compare MOLAR SOLUTION.

normal state of atom The condition in which an atom is at its lowest energy level. For the hydrogen atom, for example, the state in which the electron is in the lowest-energy orbit.

normal-through A feature in an audio PATCH BAY or PATCH PANEL that connects two sockets by default. The top socket and the one immediately below it are connected, even when a patch cord is not plugged into either of them.

northern lights See AURORA.

north magnetic pole The north pole of the equivalent bar magnet constituted by the EARTH'S MAGNETIC FIELD. The north magnetic pole lies close to the geographic north pole. Compare SOUTH MAGNETIC POLE.

north pole **1.** See NORTH MAGNETIC POLE. **2.** The earth's geographic north pole. **3.** See NORTH-SEEKING POLE.

north-seeking pole Symbol, N. The so-called north pole of a magnet. When the magnet is suspended horizontally, this pole points in the direction of the earth's north magnetic pole. Compare SOUTH-SEEKING POLE.

Norton's equivalent An equivalent circuit based on NORTON'S THEOREM, replacing a Thevenin equivalent for a current-actuated device, such as a bipolar transistor. Also see THEVENIN'S THEOREM.

Norton's theorem With reference to a particular set of terminals, any network containing any number of generators and any number of constant impedances can be simplified to one constant-current generator and one impedance. The equivalent circuit will deliver to a given load the same current that would flow if the output terminals of the original circuit where short-circuited. Compare COMPENSATION THEOREM, MAXIMUM POWER TRANSFER THEOREM, RECIPROCITY THEOREM, SUPERPOSITION THEOREM, and THEVENIN'S THEOREM.

NOT In binary logic, an operation that changes high to low and vice-versa. Also see NAND CIRCUIT, NOR CIRCUIT, NOR GATE, NOT CIRCUIT, and NOT-OR CIRCUIT.

NOT-AND circuit See NAND CIRCUIT.

notation The way that numbers, quantities, or formulas are represented (e.g., *binary notation*, *Polish notation*, and *scientific notation*).

notch A dip in frequency response, typical of a band-suppression (band-elimination) filter or other frequency-rejection circuit. Compare PEAK, **3**.

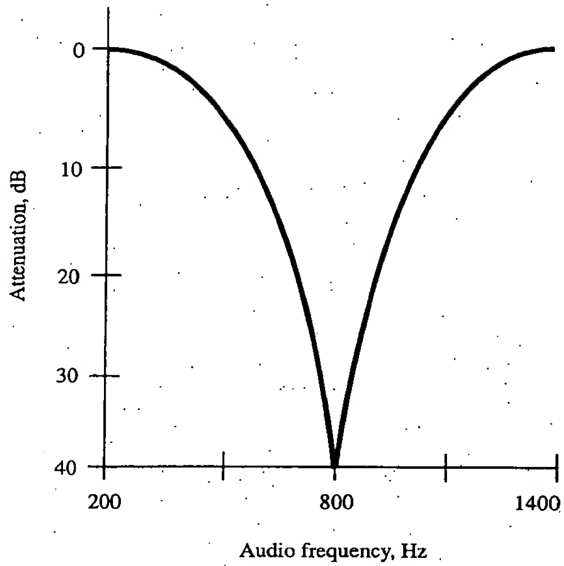
notch amplifier An amplifier containing a notch filter or other arrangement that permits it to reject one frequency or a given band of frequencies while passing all higher and lower frequencies.

notch antenna An antenna with a slot in the radiating surface, for the purpose of obtaining a directional response.

notcher See NOTCH FILTER.

notcher-peaker A circuit or device that can be set to perform either as a NOTCH FILTER or PEAK FILTER.

notch filter A frequency-rejection circuit, such as a band-suppression filter, for producing a notch.



notch

notch gate In radar, a gate that determines the minimum and maximum range.

notch sweep An oscilloscope sweep that expands only a small portion (notch) of the pattern on the screen, leaving the portions on either side of the notch untouched. Thus, the first dozen or so cycles might appear at the normal sweep speed, the next two cycles expanded, and the remaining two or three at normal sweep speed.

NOT circuit A logic circuit that provides an output pulse when there is no input pulse, and vice-versa. Also called COMPLEMENTER, NEGATOR, and INVERTER.

note See BEAT NOTE.

notebook computer A portable personal computer, also called a *laptop computer.* It is about the size of a typical three-ring notebook, and generally contains a DISKETTE DRIVE, a HARD DISK, a MODEM, and attachments for peripherals, such as printers. It uses rechargeable batteries and can be operated for approximately two to six hours between battery charges.

NOT gate A digital circuit that inverts a logical condition—either from high (logic 1) to low (logic 0) or vice-versa. Also called an *inverter.*

NOT-OR circuit A logical OR CIRCUIT combined with a NOT CIRCUIT.

novelty calculator See SPECIAL-PURPOSE CALCULATOR.

November Phonetic alphabet code word for the letter N.

novice **1.** A beginner class of amateur radio license. **2.** Any beginner or inexperienced practitioner.

no-voltage release In the starting box for a shunt motor, the electromagnet that normally holds the arm in full-running position. It is connected directly across the power line to disconnect the motor in the event of power failure. When the arm is released, it falls to its off position, thereby preventing burnout that would result if the motor were left connected to the line in the full-running position when power resumed. Compare NO-FIELD RELEASE.

noys scale A scale of apparent acoustic noise, based on a linear function instead of the more common logarithmic function.

Np **1.** Symbol for NEPTUNIUM. **2.** Abbreviation of NEPER.

N_p Symbol for number of primary turns in a transformer.

n-phase system A polyphase system having n phases.

npin transistor A junction transistor having an intrinsic layer between a p-type base and an n-type collector. The emitter is a second n-type layer on the other side of the base.

N plant See NUCLEAR POWER PLANT.

n-plus-one address instruction A computer program instruction containing two addresses, one of which specifies the location of an upcoming instruction to be executed.

NPM Symbol for *counts per minute.*

npnp device A semiconductor switching device having three junctions. Examples: FOUR-LAYER DIODE, and SILICON-CONTROLLED RECTIFIER. Also called *pnpn device.*

npn transistor A bipolar transistor in which the emitter and collector layers are n-type semiconductor material, and the base layer is p-type semiconductor material. Compare PNP TRANSISTOR.

NPO Abbreviation of NEGATIVE POSITIVE ZERO.

NPO capacitor A fixed capacitor exhibiting temperature-compensating ability over a wide temperature range, in which the coefficient has negative, positive, and zero values.

NPS Symbol for *counts per second.*

N radiation X rays emitted as a result of an electron becoming an N electron.

NRD Abbreviation of NEGATIVE-RESISTANCE DIODE.

N region See N LAYER.

NRZ Abbreviation of NONRETURN TO ZERO.

N_s Symbol for number of secondary turns in a transformer.

ns Abbreviation of NANOSECOND.

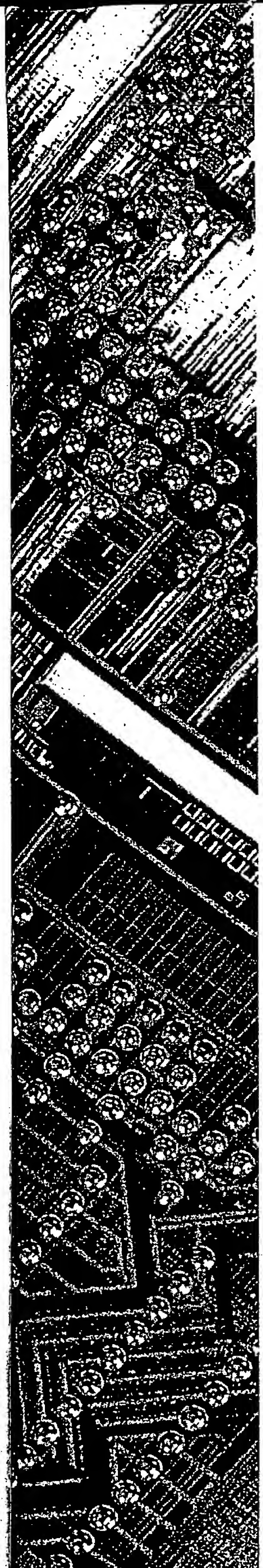
N scan See N DISPLAY.

N scope Colloquialism for a radar set using an N DISPLAY.

nsec Alternate abbreviation of NANOSECOND.

Ns/m² Newton-seconds per meter squared, the unit of dynamic viscosity.

n-space A coordinate system in n variables. It is generally of mathematical interest. The coordinates are written $(x_1, x_2, x_3, \ldots, x_n)$ and are called *ordered n-tuples.*



THE CIRCUITS and FILTERS HANDBOOK

Editor-in-Chief

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XIII

Filter Characteristics

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65

General Characteristics of Filters

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University of Victoria, Canada

65.1 Introduction

An electrical filter is a system that can be used to modify, reshape, or manipulate the frequency spectrum of an electrical signal according to some prescribed requirements. For example, a filter may be used to amplify or attenuate a range of frequency components, reject or isolate one specific frequency component, and so on. The applications of electrical filters are numerous, for example,

- to eliminate signal contamination such as noise in communication systems
- to separate relevant from irrelevant frequency components
- to detect signals in radios and TV's
- to demodulate signals
- to bandlimit signals before sampling
- to convert sampled signals into continuous-time signals
- to improve the quality of audio equipment, e.g., loudspeakers
- in time-division to frequency-division muliplex systems

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where zeros and poles occur in complex conjugate pairs, i.e., $z_2 = z_1^*$ and $p_2 = p_1^*$. Such a circuit is commonly referred to as a *biquad.*

After some manipulation, the transfer function in (65.26) can be expressed as

$$\begin{aligned} H_{BQ}(s) &= K \frac{s^2 + (2 \operatorname{Re} z_1)s + (\operatorname{Re} z_1)^2 + (\operatorname{Im} z_1)^2}{s^2 + (2 \operatorname{Re} p_1)s + (\operatorname{Re} p_1)^2 + (\operatorname{Im} p_1)^2} \\ &= K \frac{s^2 + (\omega_z/Q_z)s + \omega_z^2}{s^2 + (\omega_p/Q_p)s + \omega_p^2} \end{aligned}$$

where $K = a_2$, ω_z, and ω_p are the zero and pole frequencies, and Q_z and Q_p are the zero and pole *quality factors* (or Q factors for short), respectively. The formulas for the various parameters are as follows:

$$\begin{aligned} \omega_z &= \sqrt{(\operatorname{Re} z_1)^2 + (\operatorname{Im} z_1)^2} \\ \omega_p &= \sqrt{(\operatorname{Re} p_1)^2 + (\operatorname{Im} p_1)^2} \\ Q_z &= \frac{\omega_z}{2 \operatorname{Re} z_1} \\ Q_p &= \frac{\omega_p}{2 \operatorname{Re} p_1} \end{aligned}$$

The zero and pole frequencies are approximately equal to the frequencies of minimum gain and maximum gain, respectively. The zero and pole Q factors have to do with the selectivity of the filter. A high zero Q factor results in a deep notch in the amplitude response, whereas a high pole Q factor results in a very peaky amplitude response.

The dc gain and the gain as $\omega \to \infty$ in dB are given by

$$M_0 = 20 \log |H_{BQ}(0)| = 20 \log\left(K \frac{\omega_z^2}{\omega_p^2}\right)$$

and

$$M_\infty = 20 \log |H_{BQ}(j\infty)| = 20 \log K$$

respectively.

Types of Basic Filter Sections

Depending on the values of the transfer function coefficients, five basic types of filter sections can be identified, namely, low-pass, high-pass, bandpass, notch (sometimes referred to as bandreject), and allpass. These sections can serve as building blocks for the design of filters that can satisfy arbitrary specifications. They are actually sufficient for the design of all the standard types of filters, namely, Butterworth, Chebyshev, inverse-Chebyshev, and elliptic filters.

Low-pass Section

In a *low-pass* section, we have $a_2 = a_1 = 0$ and $a_0 = K\omega_p^2$. Hence the transfer function assumes the form

$$H_{LP}(s) = \frac{a_0}{s^2 + b_1 s + b_0} = \frac{K\omega_p^2}{s^2 + (\omega_p/Q_p)s + \omega_p^2}$$

[See Fig. 65.11(a).]

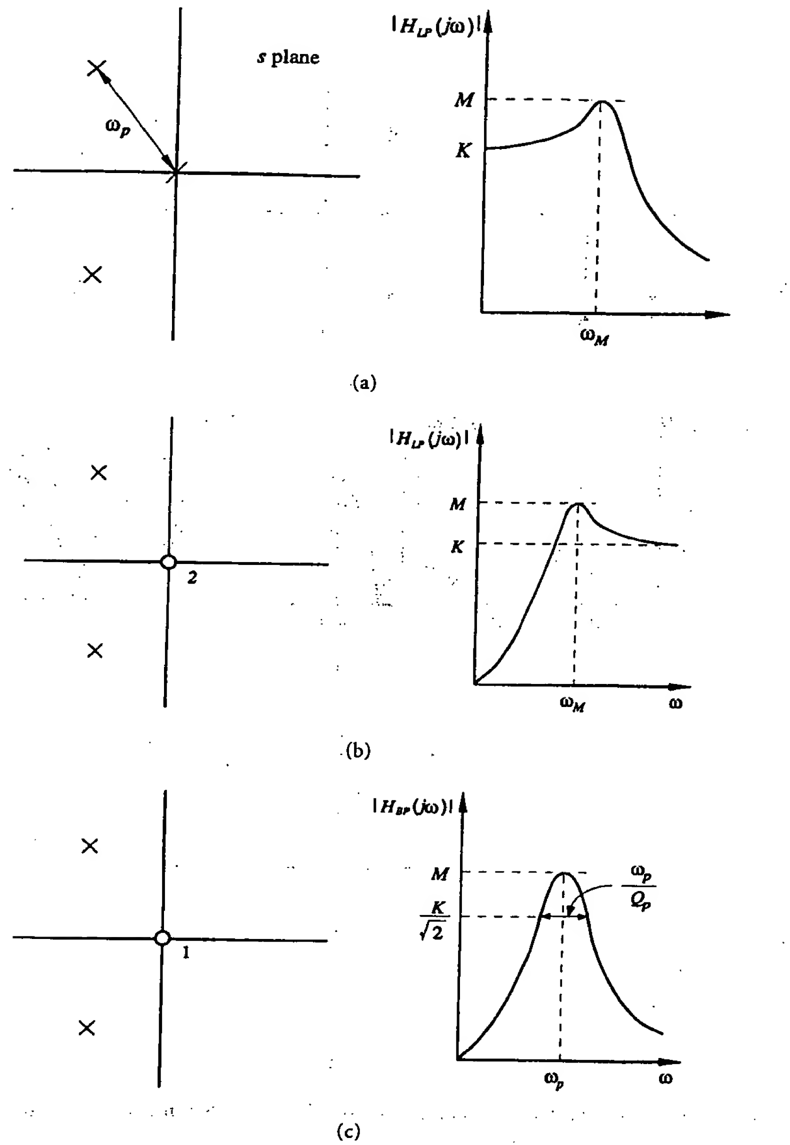
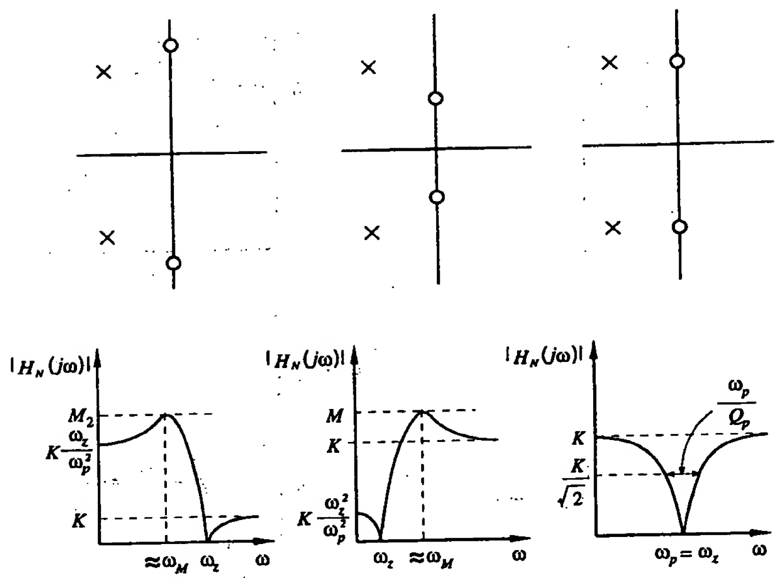
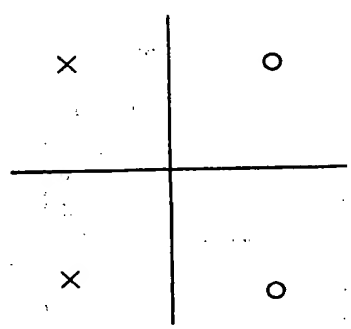


FIGURE 65.11 Basic second-order filter sections: (a) low-pass, (b) high-pass, (c) bandpass.



(d)



(e)

FIGURE 65.11 (Continued) Basic second-order filter sections: (d) notch, (e) allpass.

High-Pass Section

In a *high-pass* section, we have $a_2 = K$ and $a_1 = a_0 = 0$. Hence the transfer function assumes the form

$$H_{\mathrm{HP}}(s) = \frac{a_2 s^2}{s^2 + b_1 s + b_0} = \frac{K s^2}{s^2 + (\omega_p/Q_p)s + \omega_p^2}$$

[See Fig. 65.11(b).]

Bandpass Section

In a *bandpass* section, we have $a_1 = K\omega_p/Q_p$ and $a_2 = a_0 = 0$. Hence the transfer function assumes the form

$$H_{\mathrm{BP}}(s) = \frac{a_1 s}{s^2 + b_1 s + b_0} = \frac{K(\omega_p/Q_p)s}{s^2 + (\omega_p/Q_p)s + \omega_p^2}$$

[See Fig. 65.11(c).]

Notch Section

In a *notch* section, we have $a_2 = K$, $a_1 = 0$, and $a_0 = K\omega_z^2$. Hence the transfer function assumes the form

$$H_{\mathrm{N}}(s) = \frac{a_2 s^2 + a_0}{s^2 + b_1 s + b_0} = \frac{K(s^2 + \omega_z^2)}{s^2 + (\omega_p/Q_p)s + \omega_p^2}$$

[See Fig. 65.11(d).]

Allpass Section

In an *allpass* section, we have $a_2 = K$, $a_1 = -K\omega_p/Q_p$, and $a_0 = K\omega_p^2$. Hence the transfer function assumes the form

$$H_{\mathrm{AP}}(s) = \frac{a_2 s^2 + a_1 s + a_0}{s^2 + b_1 s + b_0} = \frac{K[s^2 - (\omega_p/Q_p)s + \omega_p^2]}{s^2 + (\omega_p/Q_p)s + \omega_p^2}$$

[See Fig. 65.11(e).]

The design of active and switched-capacitor filters is treated in some detail in Section XV.

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ELECTRONIC FILTER DESIGN HANDBOOK

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To My Family My Wife Ellen
and Children Howard, Bonnie, and Robin
Mrs. Jean Williams and Mr. and Mrs. Marcus Fuhr
for all their Love, Encouragement, and Inspiration

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A center-tapped inductor is not always available or practical. An alternate form of a bridged-T is given in figure 6-15. The parallel resonant trap design of figure 6-9 is modified by splitting the capacitor into two capacitors of twice the value, and a resistor of $\omega_0 LQ_L/4$ is introduced. The two capacitors should be closely matched.

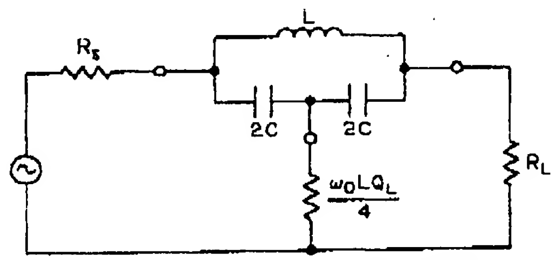


Fig. 6-15 Alternate form of bridged-T.

In conclusion, the bridged-T structure is an economical and effective means of increasing the available notch rejection of a parallel resonant trap without increasing the inductor Q. However, as a final general comment, a single null section can provide high rejection only at a single frequency or relatively narrow band of frequencies for a given 3-dB bandwidth, since $n = 1$. The stability of the circuit then becomes a significant factor. A higher-order band-reject filter design can have a wider stopband and yet maintain the same 3-dB bandwidth.

6.2 ACTIVE BAND-REJECT FILTERS

This section considers the design of active band-reject filters for both wide-band and narrow-band applications. Active null networks are covered and the popular twin-T circuit is discussed in detail.

Wide-Band Active Band-Reject Filters

Wide-band filters can be designed by first separating the specification into individual low-pass and high-pass requirements. Low-pass and high-pass filters are then independently designed and combined by paralleling the inputs and summing both outputs to form the band-reject filter.

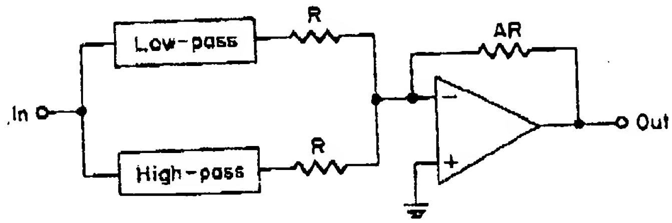
A wide-band approach is valid when the separation between cutoffs is an octave or more for all-pole filters so that minimum interaction occurs in the stopband when the outputs are summed (see section 2.1 and figure 2-13). Elliptic-function networks will require less separation, since their characteristics are steeper.

An inverting amplifier is used for summing and can also provide gain. Filters can be combined using the configuration of figure 6-16*a*, where R is arbitrary and A is the desired gain. The individual filters should have a low output impedance to avoid loading by the summing resistors.

The VCVS elliptic-function low-pass and high-pass filters of sections 3.2 and 4.2 each require an RC termination on the last stage to provide the real pole. These elements can be combined with the summing resistors, resulting in the circuit of figure 6-16*b*. R_a and C_a correspond to the denormalized values of R_5 and C_5 for the low-pass filter of figure 3-20. The denormalized high-pass

filter real-pole values are R_b and C_b. If only one filter is of the VCVS type, the summing network of the filter having the low output impedance can be r placed by a single resistor having a value of R.

When one or both filters are of the elliptic-function type, the ultimate attenuation obtainable is determined by the filter having the lesser value of A_{min}, since the stopband output is the summation of the contributions of both filters.



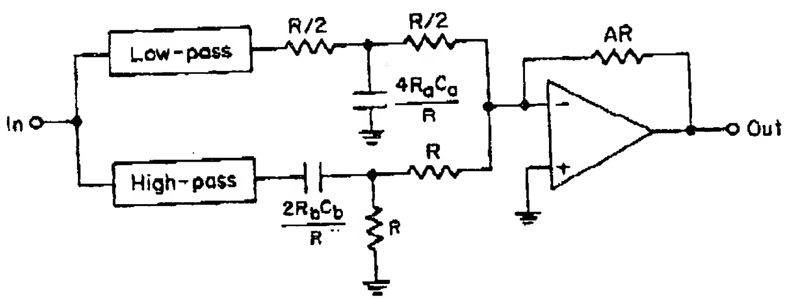


Fig. 6-16 Wide-band band-reject filters: *(a)* combining of filters having low output impedance; *(b)* combined filters requiring *RC* real poles.

Example 6-6

REQUIRED: Design an active band-reject filter having 3-dB points at 100 and 400 Hz and greater than 35 dB of attenuation between 175 and 225 Hz.

RESULT: *(a)* Since the ratio of upper cutoff to lower cutoff is well in excess of an octave, a wide-band approach can be used. First separate the specification into individual low-pass and high-pass requirements.

Low-pass:	**High-pass:**
3 dB at 100 Hz	3 dB at 400 Hz
35 dB minimum at 175 Hz	35 dB minimum at 225 Hz

(b) The low-pass and high-pass filters can now be independently designed as follows:

Low-pass filter:

Compute the steepness factor.

$$A_s = \frac{f_s}{f_c} = \frac{175\ \text{Hz}}{100\ \text{Hz}} = 1.75 \qquad (2\text{-}11)$$

An $n = 5$ Chebyshev filter having a 0.5-dB ripple is chosen using figure 2-44. The normalized active low-pass filter values are given in tabl 12-39 and the circuit is shown in figure 6-17*a*.

To denormalize the filter, multiply all resistors by Z and divide all capacitors by $Z \times$ FSF, where Z is conveniently selected at 10^4 and the FSF is $2\pi f_c$, where f_c is 100 Hz. The denormalized low-pass filter is given in figure 6-17*b*.

High-pass filter:

Compute the steepness factor.

$$A_s = \frac{f_c}{f_s} = \frac{400 \text{ Hz}}{225 \text{ Hz}} = 1.78 \qquad (2\text{-}13)$$

An $n = 5$ Chebyshev filter with a 0.5-dB ripple will also satisfy the high-pass requirement. A high-pass transformation can be performed on the normalized low-pass filter of figure 6-17*a* to obtain the circuit of figure 6-17*c*. All resistors have been replaced with capacitors and vice versa using reciprocal element values.

The normalized high-pass filter is then frequency- and impedance-scaled by multiplying all resistors by Z and dividing all capacitors by $Z \times$ FSF, where Z is chosen at 10^5 and FSF is $2\pi f_c$, using an f_c of 400 Hz. The denormalized high-pass filter is shown in figure 6-17*d* using standard 1% resistor values.

(*c*) The individual low-pass and high-pass filters can now be combined using the configuration of figure 6-16*a*. Since no gain is required, A is set equal to unity. The value of R is conveniently selected at 10 kΩ, resulting in the circuit of figure 6-17*e*.

Band-Reject Transformation of Low-Pass Poles

The wide-band approach to the design of band-reject filters using combined low-pass and high-pass networks is applicable to bandwidths of typically an octave or more. If the separation between cutoffs is insufficient, interaction in the stopband will occur, resulting in inadequate stopband rejection (see figure 2-13).

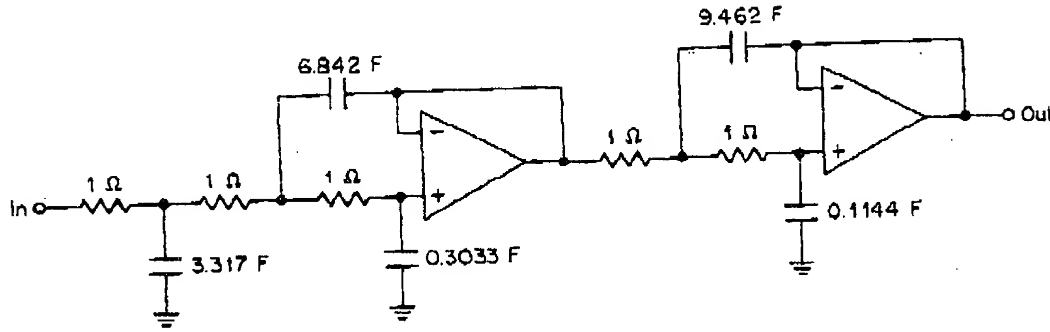
A more general approach involves normalizing the band-reject requirement and selecting a normalized low-pass filter type that meets these specifications. The corresponding normalized low-pass poles are then directly transformed to the band-reject form and realized using active sections.

A band-reject transfer function can be derived from a low-pass transfer function by substituting the frequency variable f by a new variable given by

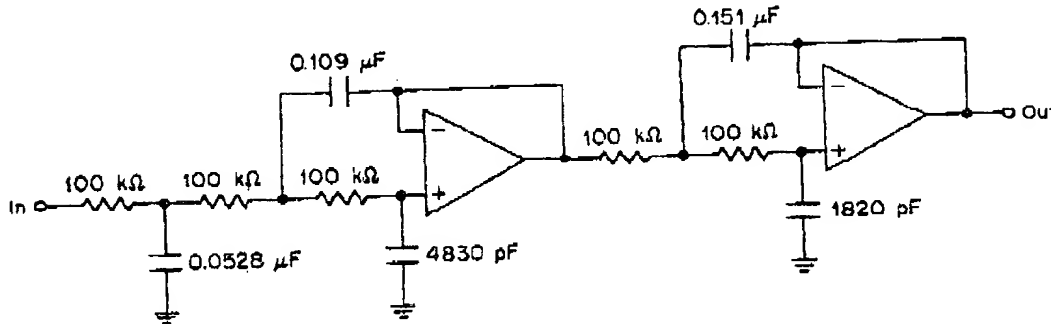
$$f_{\text{br}} = \frac{1}{f_0\left(\dfrac{f}{f_0} - \dfrac{f_0}{f}\right)} \qquad (6\text{-}32)$$

This transformation combines the low-pass to high-pass and subsequent band-reject transformation discussed in section 6.1 so that a band-reject filter can be obtained directly from the low-pass transfer function.

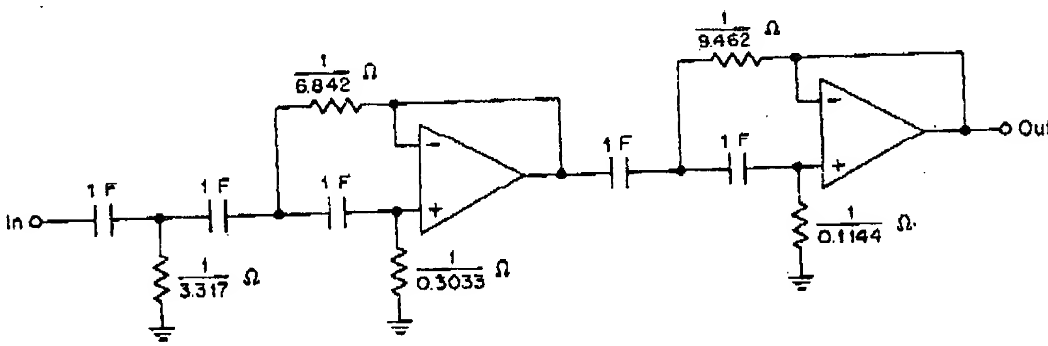
The band-reject transformation results in two pairs of complex poles and a pair of second-order imaginary zeros from each low-pass complex pole pair. A single low-pass real pole is transformed into a complex pole pair and a pair of first-order imaginary zeros. These relationships are illustrated in figure 6-18. The zeros occur at center frequency and result from the transformed low-pass zeros at infinity.


9.462 F
6.842 F
1 Ω
1 Ω
Out
1 Ω
1 Ω
1 Ω
In
0.1144 F
3.317 F
0.3033 F

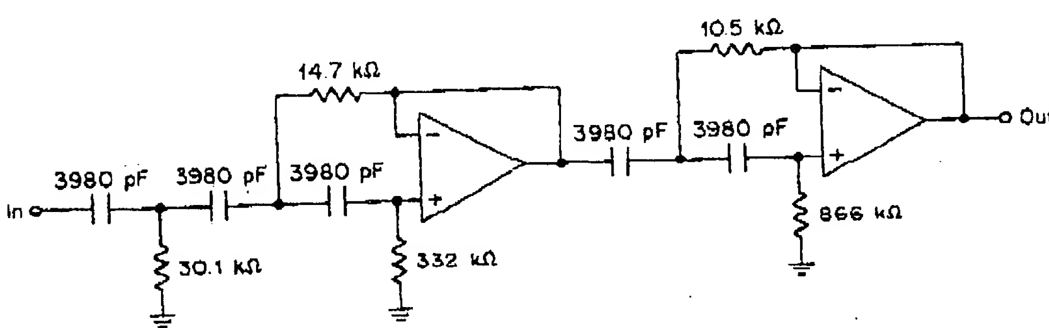
(a)


0.151 μF
0.109 μF
Out
100 kΩ
100 kΩ
100 kΩ
100 kΩ
100 kΩ
In
1820 pF
0.0528 μF
4830 pF

(b)


1/9.462 Ω
1/6.842 Ω
Out
1 F
1 F
1 F
1 F
1 F
In
1/0.1144 Ω
1/3.317 Ω
1/0.3033 Ω

(c)


10.5 kΩ
14.7 kΩ
Out
3980 pF
3980 pF
3980 pF
3980 pF
3980 pF
In
866 kΩ
30.1 kΩ
332 kΩ

(d)

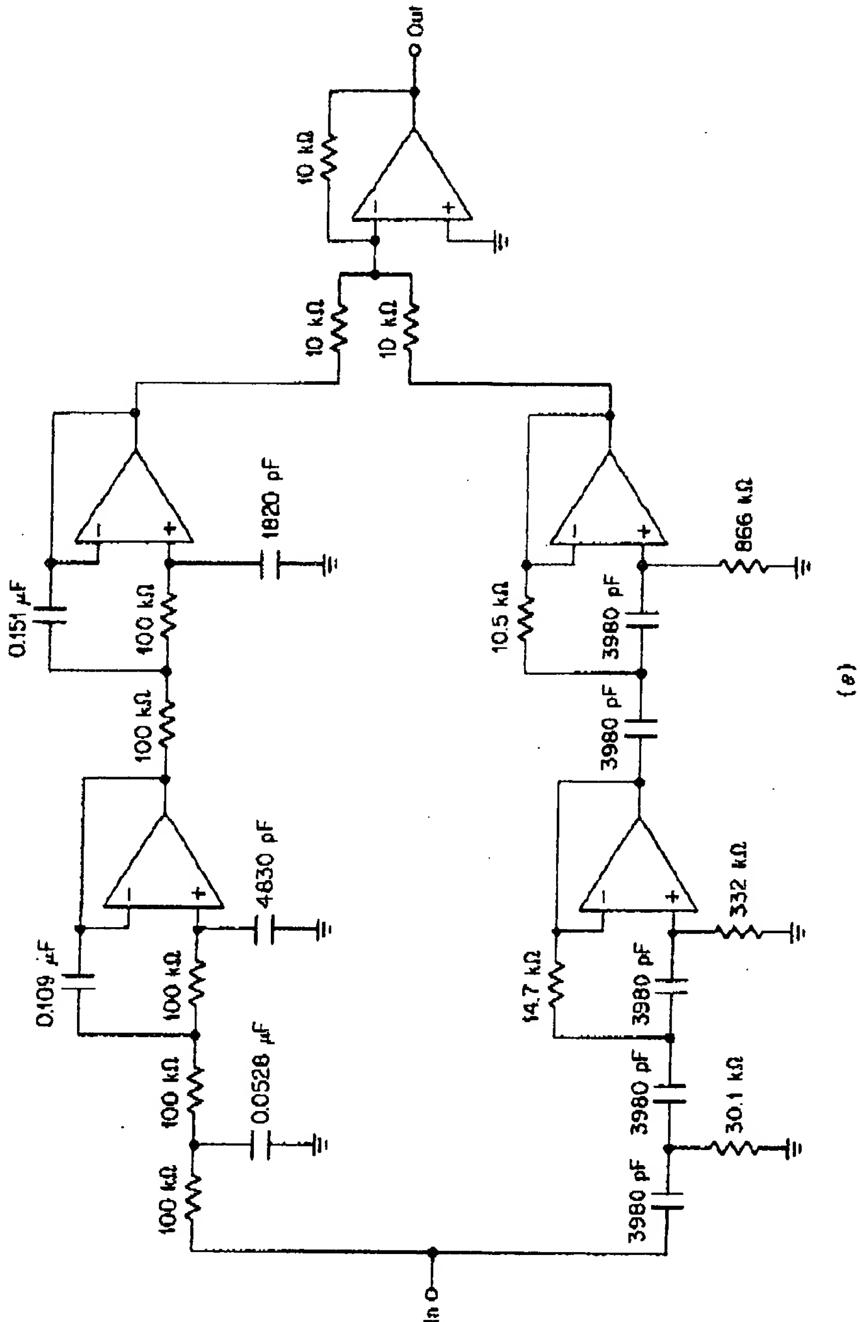


Fig. 6-17 Wide-band band-reject filter of example 6-6: *(a)* normalized low-pass filter; *(b)* denormalized low-pass filter; *(c)* transformed normalized high-pass filter; *(d)* denormalized high-pass filter; *(e)* combining filters to obtain band-reject response.

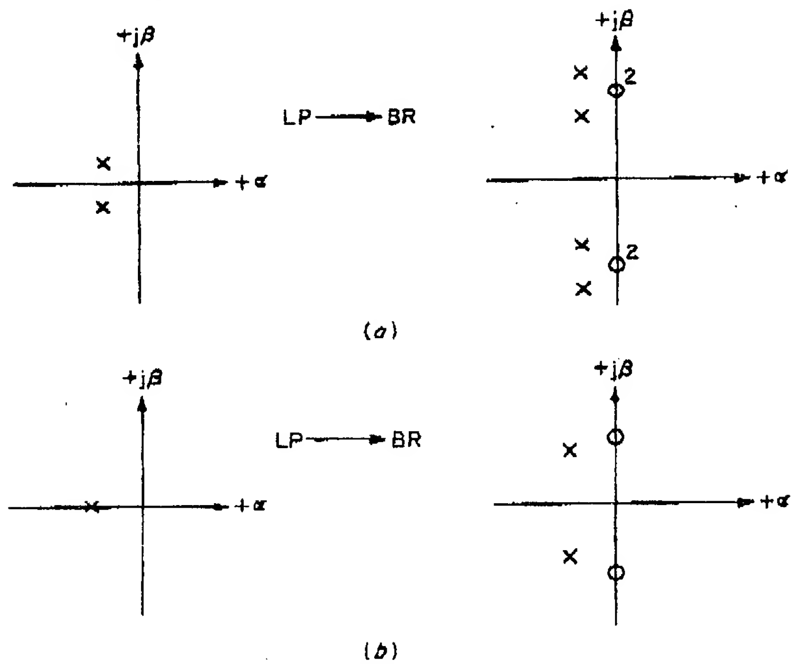


Fig. 6-18 Band-reject transformation of low-pass poles: *(a)* low-pass complex pole pair; *(b)* low-pass real pole.

The band-reject pole-zero pattern of figure 6-18*a* corresponds to two band-reject sections where each section provides a zero at center frequency and provides one of the pole pairs. The pattern of figure 6-18*b* is realized by a single band-reject section where the zero also occurs at center frequency.

To make the low-pass to band-reject transformation, first compute

$$Q_{br} = \frac{f_0}{\mathrm{BW}} \tag{6-33}$$

where f_0 is the geometric center frequency and BW is the passband bandwidth. The transformation then proceeds as follows:

Complex Poles The tables of chapter 12 contain tabulated poles corresponding to the all-pole low-pass filter families discussed in chapter 2. Complex poles are given in the form: $-\alpha \pm j\beta$ where α is the real coordinate and β is the imaginary part. Given α, β, Q_{br}, and f_0, the following computations result in two sets of values for Q and frequency which defines two band-reject filter sections. Each section also has a zero at f_0.

$$C = \alpha^2 + \beta^2 \tag{6-34}$$

$$D = \frac{\alpha}{Q_{br} C} \tag{6-35}$$

$$E = \frac{\beta}{Q_{br} C} \tag{6-36}$$

$$F = E^2 - D^2 + 4 \tag{6-37}$$

$$G = \sqrt{\frac{F}{2} + \sqrt{\frac{F^2}{4} + D^2E^2}} \tag{6-38}$$

$$H = \frac{DE}{G} \tag{6-39}$$

$$K = \frac{1}{2}\sqrt{(D+H)^2 + (E+G)^2} \tag{6-40}$$

$$Q = \frac{K}{D+H} \tag{6-41}$$

$$f_{ra} = \frac{f_0}{K} \tag{6-42}$$

$$f_{rb} = K f_0 \tag{6-43}$$

$$f_\infty = f_0 \tag{6-44}$$

The two band-reject sections have resonant frequencies of f_{ra} and f_{rb} (in hertz) and identical Q's given by equation (6-41). In addition each section has a zero at f_0, the filter geometric center frequency.

Real Poles A normalized low-pass real pole having a real coordinate of α_0 is transformed into a single band-reject section having a Q given by

$$Q = Q_{br}\,\alpha_0 \tag{6-45}$$

The section resonant frequency is equal to f_0, and the section must also have a transmission zero at f_0.

Example 6-7

REQUIRED: Determine the pole and zero locations for a band-reject filter having the following specifications:

Center frequency of 3600 Hz
3 dB at ±150 Hz
40 dB minimum at ±30 Hz

RESULT: (*a*) Since the filter is narrow, the requirement can be treated directly in its arithmetically symmetrical form:

$$f_0 = 3600 \text{ Hz}$$
$$BW_{3\text{ dB}} = 300 \text{ Hz}$$
$$BW_{40\text{ dB}} = 60 \text{ Hz}$$

The band-reject steepness factor is given by

$$A_s = \frac{\text{passband bandwidth}}{\text{stopband bandwidth}} = \frac{300 \text{ Hz}}{60 \text{ Hz}} = 5 \tag{2-20}$$

(*b*) An $n = 3$ Chebyshev normalized low-pass filter having a 0.1-dB ripple is selected using figure 2-42. The corresponding pole locations are found in table 12-28 and are

$-0.3500 \pm j0.8695$
-0.6999

First make the preliminary computation:

$$Q_{br} = \frac{f_0}{BW_{3\text{ dB}}} = \frac{3600 \text{ Hz}}{300 \text{ Hz}} = 12 \tag{6-33}$$

6-26 Band-Reject Filters

The low-pass to band-reject pole transformation is performed as follows:

Complex-pole transformation:

$$\alpha = 0.3500 \qquad \beta = 0.8695$$

$$C = \alpha^2 + \beta^2 = 0.878530 \tag{6-34}$$

$$D = \frac{\alpha}{Q_{br}C} = 0.033199 \tag{6-35}$$

$$E = \frac{\beta}{Q_{br}C} = 0.082477 \tag{6-36}$$

$$F = E^2 - D^2 + 4 = 4.005700 \tag{6-37}$$

$$G = \sqrt{\frac{F}{2} + \sqrt{\frac{F^2}{4} + D^2E^2}} = 2.001425 \tag{6-38}$$

$$H = \frac{DE}{G} = 0.001368 \tag{6-39}$$

$$K = \frac{1}{2}\sqrt{(D+H)^2 + (E+G)^2} = 1.042094 \tag{6-40}$$

$$Q = \frac{K}{D+H} = 30.15 \tag{6-41}$$

$$f_{ra} = \frac{f_0}{K} = 3455 \text{ Hz} \tag{6-42}$$

$$f_{rb} = Kf_0 = 3752 \text{ Hz} \tag{6-43}$$

$$f_\infty = f_0 = 3600 \text{ Hz} \tag{6-44}$$

Real-pole transformation:

$$\alpha_0 = 0.6999$$

$$Q = Q_{br}\,\alpha_0 = 8.40 \tag{6-45}$$

$$f_r = f_\infty = f_0 = 3600 \text{ Hz}$$

The block diagram is shown in figure 6-19.

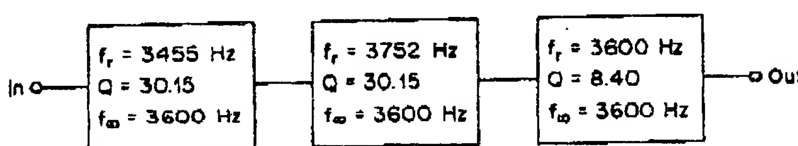


Fig. 6-19 Block diagram of example 6-7.

Narrow-Band Active Band-Reject Filters

Narrow-band active band-reject filters are designed by first transforming a set of normalized low-pass poles to the band-reject form. The band-reject poles are computed in terms of resonant frequency f_r, Q, and f_∞ using the results of section 6.2 and are then realized with active band-reject sections.

The VCVS Band-Reject Section Complex low-pass poles result in a set of band-reject parameters where f_r and f_∞ do not occur at the same frequency. Band-reject sections are then required that permit independent selection of f_r

and f_∞ in their design procedure. Both the VCVS and biquad circuits covered in section 5.2 under Elliptic-Function Bandpass Filters have this degree of freedom.

The VCVS realization is shown in figure 6-20. The design equations were given in section 5.2 under Elliptic-Function Bandpass Filters and are repeated

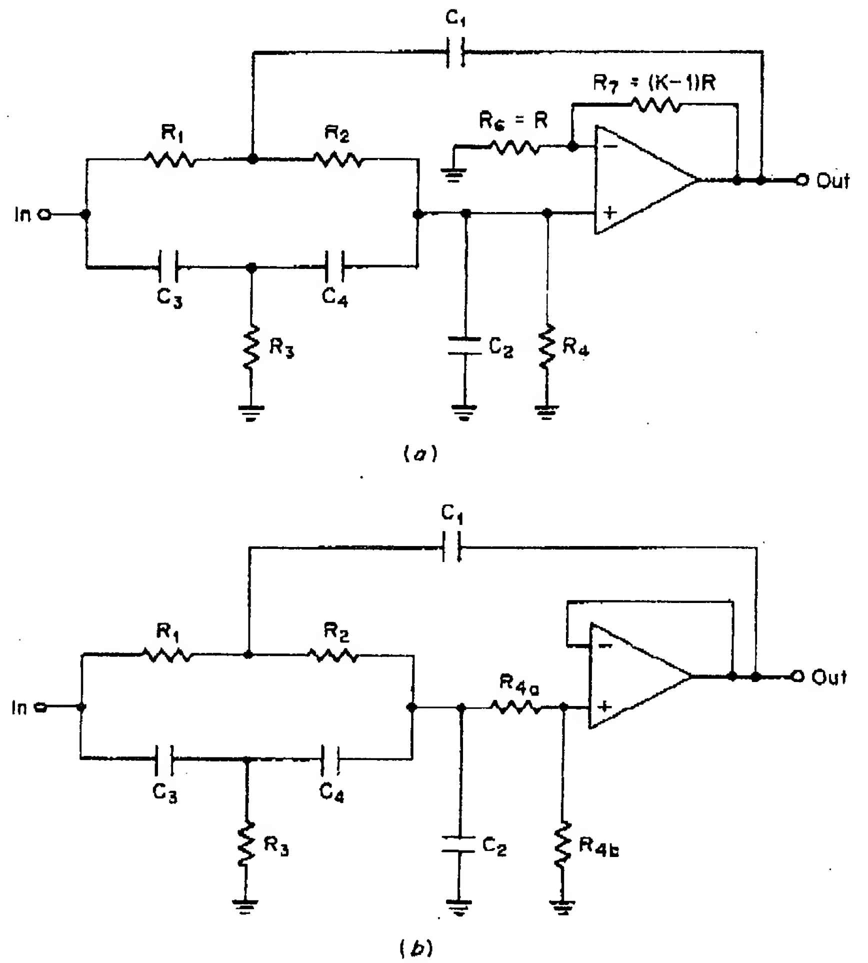


Fig. 6-20 VCVS realization for band-reject filters: *(a)* circuit for $K > 1$; *(b)* circuit for $K < 1$.

here for convenience, where f_r, Q, and f_∞ are obtained by the band-reject transformation procedure of section 6.2. The values are computed from

$$R_2 = 2R_1 \tag{6-47}$$

$$R_3 = \frac{f_\infty^2 + f_r^2}{4.5 f_r^2} R' \tag{6-48}$$

$$R_4 = 4.5R_3 \tag{6-49}$$

$$C_1 = \frac{1.5}{2\pi f_r R_1} \tag{6-50}$$

$$C_2 = \frac{C_1}{4.5} \tag{6-51}$$

$$C_3 = \frac{f_r}{2\pi R_1 f_\infty^2} \tag{6-52}$$

$$C_4 = \frac{C_3}{2} \tag{6-53}$$

$$K = \frac{\left(2.5 - \frac{1}{Q}\right)\left(\frac{f_r^2}{f_\infty^2} + 1\right)}{1.5} \tag{6-54}$$

$$R_6 = R \tag{6-55}$$
$$R_7 = (K - 1)R \tag{6-56}$$

where R' and R can be arbitrarily chosen.

The circuit of figure 6-20*a* is used when $K > 1$. In the cases where $K < 1$, the configuration of figure 6-20*b* is utilized, where

$$R_{4a} = (1 - K)R_4 \tag{6-57}$$

and

$$R_{4b} = KR_4 \tag{6-58}$$

The section gain at DC is given by

$$A_{dc} = \frac{f_\infty^2}{f_r^2 + f_\infty^2} \tag{6-59}$$

The gain of the composite filter in the passband is the product of the DC gains of all the sections.

The VCVS structure has a number of undesirable characteristics. Although the circuit Q can be adjusted by making R_6 or R_7 variable when $K > 1$, the Q cannot be independently measured since the 3-dB bandwidth at the output is affected by the transmission zero. Resonant frequency f_r or the notch frequency f_∞ cannot be easily adjusted, since these parameters are determined by the interaction of a number of elements. Also the section gain is fixed by the design parameters. Another disadvantage of the circuit is that a large spread in capacitor values[1] may occur so that standard values cannot be easily used. Nevertheless the VCVS realization makes effective use of a minimum number of operational amplifiers in comparison with other implementations and is widely used. However, because of its lack of adjustment capability, its application is generally restricted to Q's below 10 and with 1% component tolerances.

The State-Variable Band-Reject Section The biquad or state-variable elliptic-function bandpass filter section discussed in section 5.2 is highly suitable for implementing band-reject transfer functions. The circuit is given in figure 6-21. By connecting resistor R_5 to either node 1 or to node 2, the notch frequency f_∞ will be located above or below the pole resonant frequency f_r.

[1] The elliptic-function configuration of the VCVS uniform capacitor structure given in section 3.2 can be used at the expense of additional sensitivity.

Section Q's of up to 200 can b obtained. The design parameters f_r, Q, and f_∞ as well as the section gain can be independently chosen, monitored, and adjusted. From the point of view of low sensitivity and maximum flexibility the biquad approach is the most desirable method of realization.

The design equations were stated in section 5.2 under Elliptic-Function Band-pass Filters and are repeated here for convenience, where f_r, Q, and f_∞ are given and the values of C, R, and R' can be arbitrarily chosen.

$$R_1 = R_4 = \frac{Q}{2\pi f_r C} \tag{6-60}$$

$$R_2 = R_3 = \frac{R_1}{Q} \tag{6-61}$$

$$R_5 = \frac{f_r^2 R}{Q|f_r^2 - f_\infty^2|} \tag{6-62}$$

for $f_\infty > f_r$:

$$R_6 = \frac{f_r^2 R}{f_\infty^2} \tag{6-63}$$

and when $f_\infty < f_r$:

$$R_6 = R \tag{6-64}$$

The value of R_6 is based on unity section gain at DC. The gain can be raised or lowered by proportionally increasing or decreasing R_6.

Resonance is adjusted by monitoring the phase shift between the section input and node 3 using a Lissajous pattern and adjusting R_3 for 180° phase shift with an input frequency of f_r.

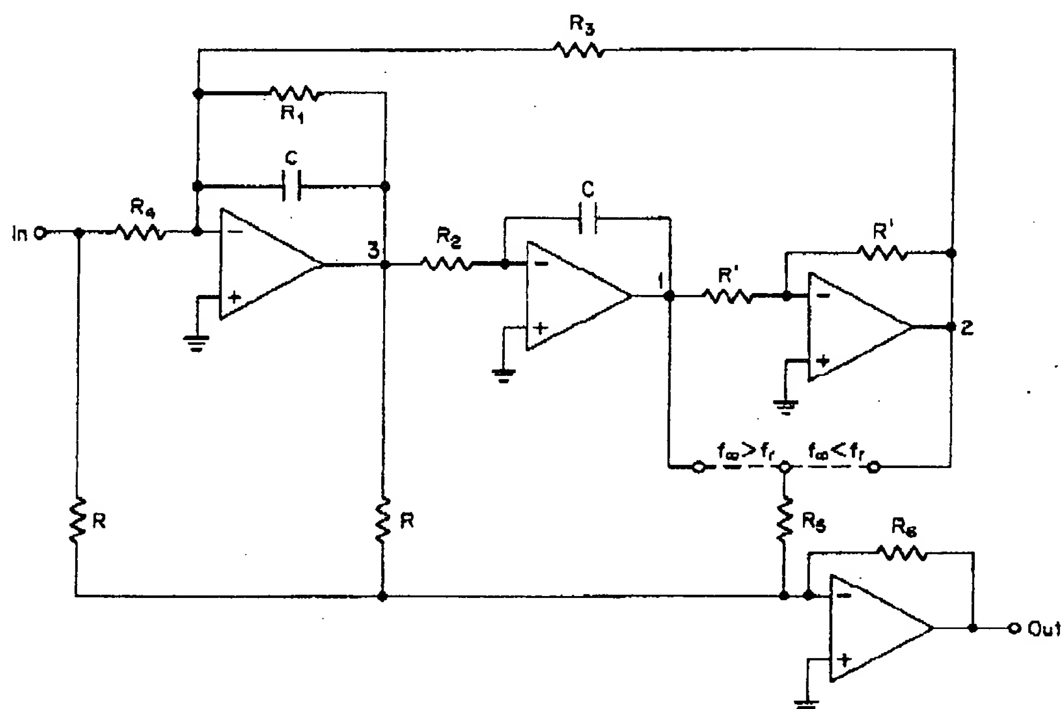


Fig. 6-21 Biquad band-reject realization.

The Q is controlled by R_1 and can be measured at node 3 in terms of section 3-dB bandwidth, or R_1 can be adjusted until unity gain occurs between the input and node 3 with f_r applied. Because of the Q enhancement effect discussed in section 5.2 under All-Pole Bandpass Configurations a Q adjustment is usually necessary.

The notch frequency is then determined by monitoring the section output for a null. Adjustment is normally not required, since tuning of f_r will usually bring in f_∞ with acceptable accuracy. If an adjustment is desired, R_5 can be made variable.

Sections for Transformed Real Poles When a real pole undergoes a band-reject transformation, the result is a single pole pair and a single set of imaginary zeros. Complex poles resulted in two sets of pole pairs and two sets of zeros. The resonant frequency f_r of the transformed real pole is exactly equal to the notch frequency f_∞; so the design flexibility of the VCVS and biquad structures is not required.

A general second-order bandpass transfer function can be expressed as

$$T(s)=\frac{\frac{\omega_r}{Q}s}{s^2+\frac{\omega_r}{Q}s+\omega_r^2} \tag{6-65}$$

where the gain is unity at ω_r. If we realize the circuit of figure 6-22 where $T(s)$ corresponds to the above transfer function, the composite transfer function at the output is given by

$$T(s)=\frac{s^2+\omega_r^2}{s^2+\frac{\omega_r}{Q}s+\omega_r^2} \tag{6-66}$$

This corresponds to a band-reject transfer function having a transmission zero at f_r (i.e., $f_\infty=f_r$). The occurrence of this zero can also be explained intuitively from the structure of figure 6-22. Since $T(s)$ is unity only at f_r, both input signals to the summing amplifier will then cancel, resulting in no output signal.

These results indicate that band-reject sections for transformed real poles can be obtained by combining any of the all-pole bandpass circuits of section 5.2 in the configuration of figure 6-22. The basic design parameters are the required f_r and Q of the band-reject section which are directly used in the design equations for the bandpass circuits.

By combining these bandpass sections with summing amplifiers, the three band-reject structures of figure 6-23 can be derived. The design equations for the bandpass sections were given in section 5.2 and are repeated here where C, R, and R' can be arbitrarily chosen.

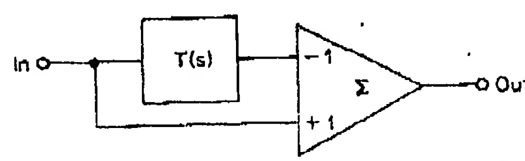


Fig. 6-22 Band-reject configuration for $f_r=f_\infty$.

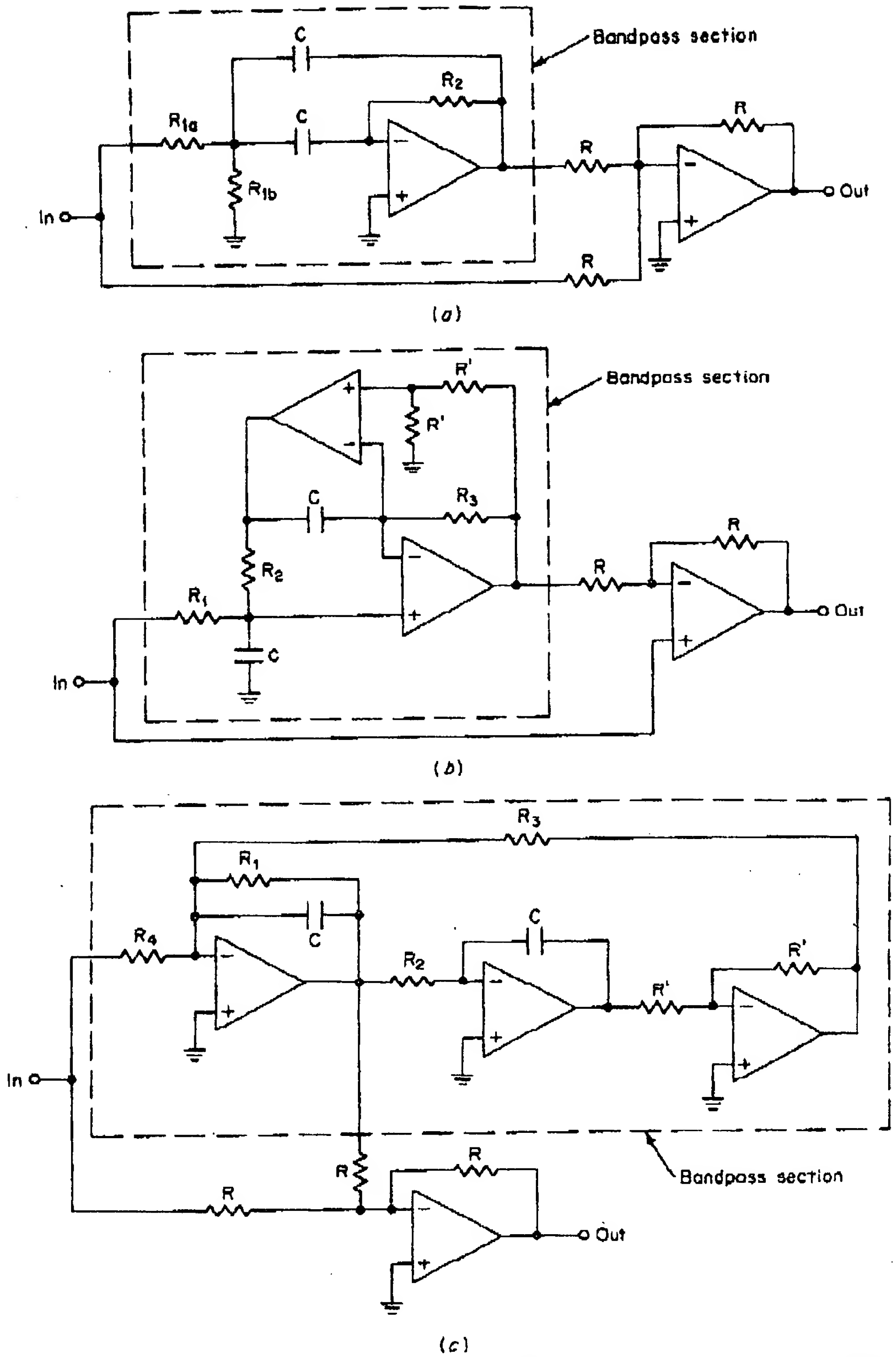


Fig. 6-23 Band-reject circuits for $f_r = f_\infty$ *(a)* MFBP band-reject section ($Q < 20$); *(b)* DABP band-reject section ($Q < 150$); *(c)* biquad band-reject section ($Q < 200$).

6-32 Band-Reject Filters

The MFBP band-reject section ($f_r = f_\infty$) is given by

$$R_2 = \frac{Q}{\pi f_r C} \quad (6\text{-}67)$$

$$R_{1a} = \frac{R_2}{2} \quad (6\text{-}68)$$

$$R_{1b} = \frac{R_{1a}}{2Q^2 - 1} \quad (6\text{-}69)$$

The DABP band-reject section ($f_r = f_\infty$) is given by

$$R_1 = \frac{Q}{2\pi f_r C} \quad (6\text{-}70)$$

$$R_2 = R_3 = \frac{R_1}{Q} \quad (6\text{-}71)$$

The biquad band-reject section ($f_r = f_\infty$) is given by

$$R_1 = R_4 = \frac{Q}{2\pi f_r C} \quad (6\text{-}72)$$

$$R_2 = R_3 = \frac{R_1}{Q} \quad (6\text{-}73)$$

These equations correspond to unity bandpass gain for the MFBP and biquad circuits so that cancellation at f_r will occur when the section input and bandpass output signals are equally combined by the summing amplifiers. Since the DABP section has a gain of 2 and has a noninverting output, the circuit of figure 6-23*b* has been modified accordingly so that cancellation occurs.

Tuning can be accomplished by making R_{1b}, R_2, and R_3 variable in the MFBP, DABP, and biquad circuits, respectively. In addition the biquad circuit will usually require R_1 to be made adjustable to compensate for the Q-enhancement effect (see section 5.2 under All-Pole Bandpass Configurations). The circuit can be tuned by adjusting the indicated elements for either a null at f_r measured at the circuit output or for 0° or 180° phase shift at f_r observed between the input and the output of the bandpass section. If the bandpass section gain is not sufficiently close to unity for the MFBP and biquad case and 2 for the DABP circuit, the null depth may be inadequate.

Example 6-8

REQUIRED: Design an active band-reject filter from the band-reject parameters determined in example 6-7 having a gain of +6 dB.

RESULT: (*a*) The band-reject transformation in example 6-7 resulted in the following set of requirements for a three-section filter:

Section	f_r	Q	f_∞
1	3455 Hz	30.15	3600 Hz
2	3752 Hz	30.15	3600 Hz
3	3600 Hz	8.40	3600 Hz

(*b*) Two biquad circuits in tandem will be used for sections 1 and 2 followed by a DABP band-reject circuit for section 3. The value

of C is chosen at 0.01 μF and R as well as R' at 10 kΩ. Since the DABP section has a gain of 2 at DC, which satisfies the 6-dB gain requirement, both biquad sections should then have unity gain. The element values are determined as follows:

Section 1 (biquad of figure 6-21):

$f_r = 3455$ Hz $\quad Q = 30.15 \quad f_\infty = 3600$ Hz

$$R_1 = R_4 = \frac{Q}{2\pi f_r C} = \frac{30.15}{2\pi \times 3455 \times 10^{-8}} = 138.9\ \text{k}\Omega \qquad (6\text{-}60)$$

$$R_2 = R_3 = \frac{R_1}{Q} = \frac{138.9 \times 10^3}{30.15} = 4610\ \Omega \qquad (6\text{-}61)$$

$$R_5 = \frac{f_r^2 R}{Q|f_r^2 - f_\infty^2|} = \frac{3455^2 \times 10^4}{30.15|3455^2 - 3600^2|} = 3870\ \Omega \qquad (6\text{-}62)$$

$$R_6 = \frac{f_r^2 R}{f_\infty^2} = \frac{3455^2 \times 10^4}{3600^2} = 9210\ \Omega \qquad (6\text{-}63)$$

Section 2 (biquad of figure 6-21):

$f_r = 3752$ Hz $\quad Q = 30.15 \quad f_\infty = 3600$ Hz

$R_1 = R_4 = 127.9$ kΩ (6-60)

$R_2 = R_3 = 4240\ \Omega$ (6-61)

$R_5 = 4180\ \Omega$ (6-62)

$R_6 = 10$ kΩ (6-64)

Section 3 (DABP of figure 6-23):

$f_r = f_\infty = 3600$ Hz $\quad Q = 8.40$

$$R_1 = \frac{Q}{2\pi f_r C} = \frac{8.40}{2\pi \times 3600 \times 10^{-8}} = 37.1\ \text{k}\Omega \qquad (6\text{-}70)$$

$$R_2 = R_3 = \frac{R_1}{Q} = \frac{37.1 \times 10^3}{8.40} = 4420\ \Omega \qquad (6\text{-}71)$$

The final circuit is shown in figure 6-24 with standard 1% resistor values. The required resistors have been made variable so that the resonant frequencies can be adjusted for all sections, and in addition the Q is variable for the biquad circuits.

Active Null Networks

Active null networks are single sections used to provide attenuation at a single frequency or over a narrow band of frequencies. The most popular sections are of the twin-T form; so this circuit will be discussed in detail along with some other structures.

The Twin-T The twin-T was first discovered by H. W. Augustadt in 1934. Although this circuit is passive by nature, it is also used in many active configurations to obtain a variety of different characteristics.

The circuit of figure 6-25*a* is an *RC* bridge structure where balance or an output null occurs at 1 rad/s when all arms have an equal impedance ($0.5 - j0.5\ \Omega$). The circuit is redrawn in the form of a symmetrical lattice in figure 6-25*b* (refer to Guillemin and Stewart in references for detailed discussions of the lattice). The lattice of figure 6-25*b* can be redrawn again in the form of two parallel lattices as shown in figure 6-25*c*.

If identical series elements are present in both the series and shunt branches of a lattice, the element may be extracted and symmetrically placed outside

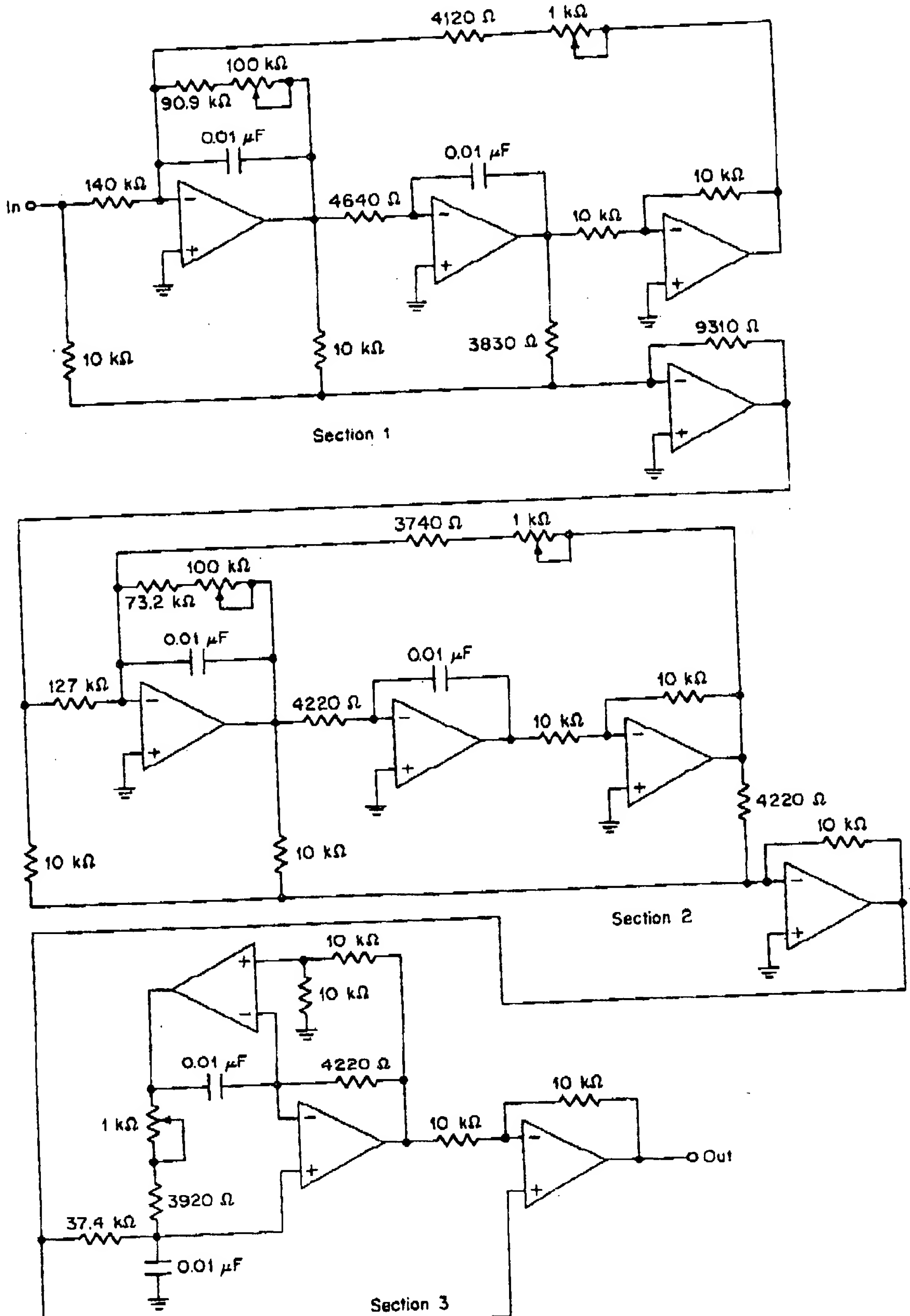


Fig. 6-24 Band-reject filter of example 6-8.

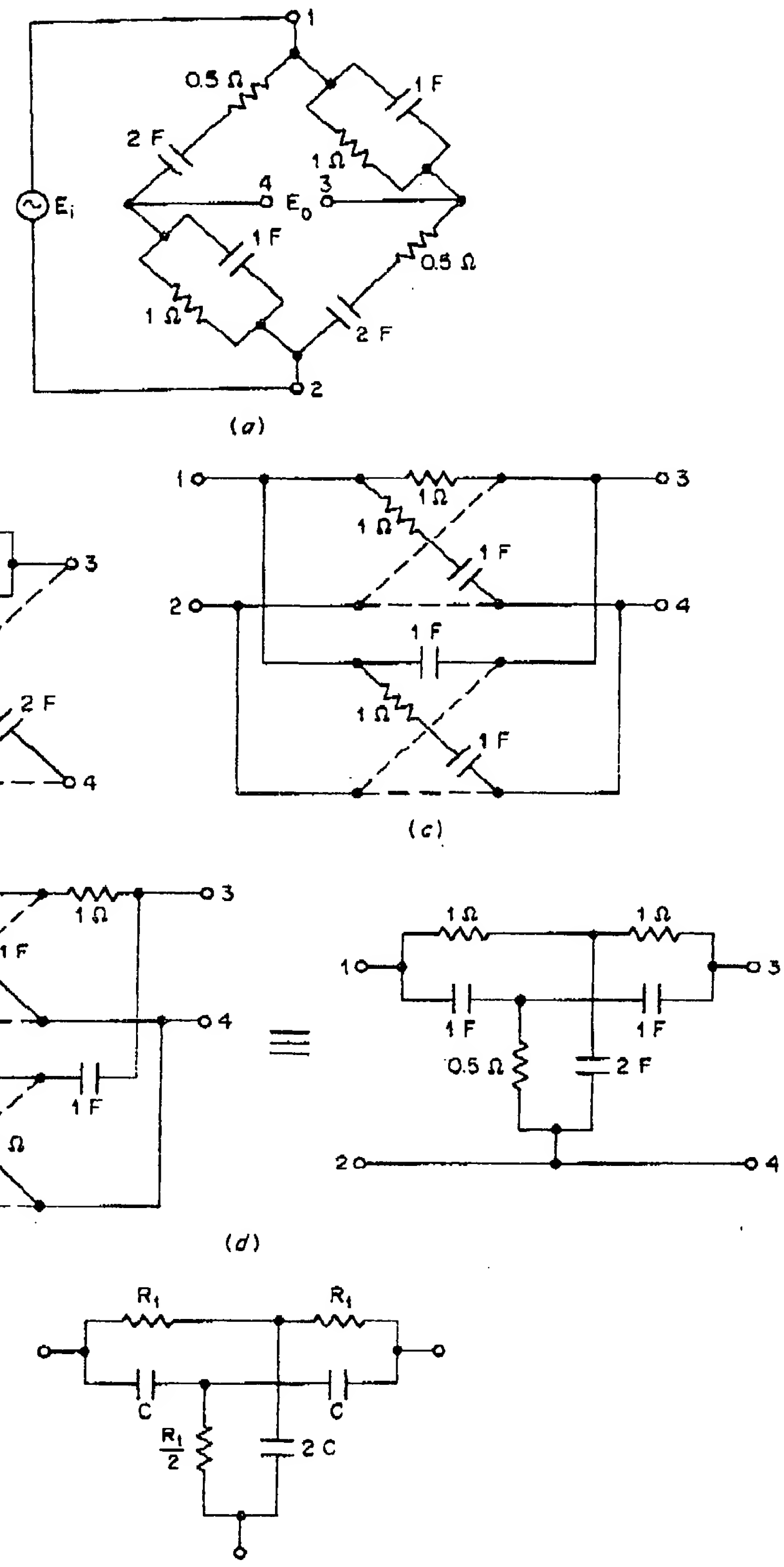


Fig. 6-25 Derivation of the twin-T: *(a) RC* bridge; *(b)* lattice circuit; *(c)* parallel lattice; *(d)* twin-T equivalent; *(e)* general form of twin-T.

the lattice structure. A 1-Ω resistor satisfies this requirement for the upper lattice and a 1-F capacitor for the lower lattice. Removal of these components to outside the lattice results in the twin-T of figure 6-25*d*.

The general form of a twin-T is shown in figure 6-25*e*. The value of R_1 is computed from

$$R_1 = \frac{1}{2\pi f_0 C} \tag{6-74}$$

where C is arbitrary. This denormalizes the circuit of figure 6-25*d* so that the null now occurs at f_0 instead of at 1 rad/s.

When a twin-T is driven from a voltage source and terminated in an infinite load,[2] the transfer function is given by

$$T(s) = \frac{s^2 + \omega_0^2}{s^2 + 4\omega_0 s + \omega_0^2} \tag{6-75}$$

If we compare this expression with the general transfer function of a second-order pole-zero section as given by equation (6-66), we can determine that a twin-T provides a notch at f_0 with a Q of ¼. The attenuation at any bandwidth can be computed by

$$A_{dB} = 10 \log \left[1 + \left(\frac{4 f_0}{BW_{x\,dB}} \right)^2 \right] \tag{6-76}$$

The frequency response is shown in figure 6-26. The requirement for geometric symmetry applies.

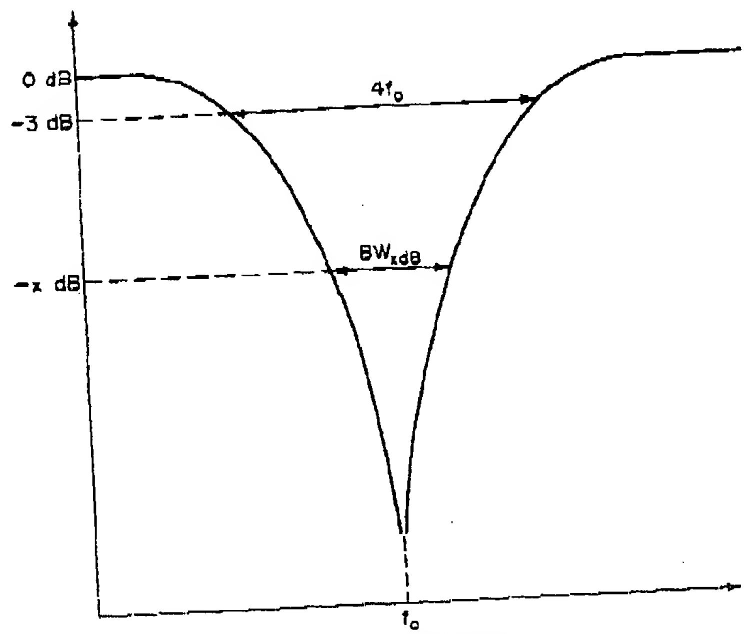


Fig. 6-26 Frequency response of twin-T.

[2] Since the source and load are always finite, the value of R_1 should be in the vicinity of $\sqrt{R_s R_L}$, provided that the ratio R_L/R_s is in excess of 10.

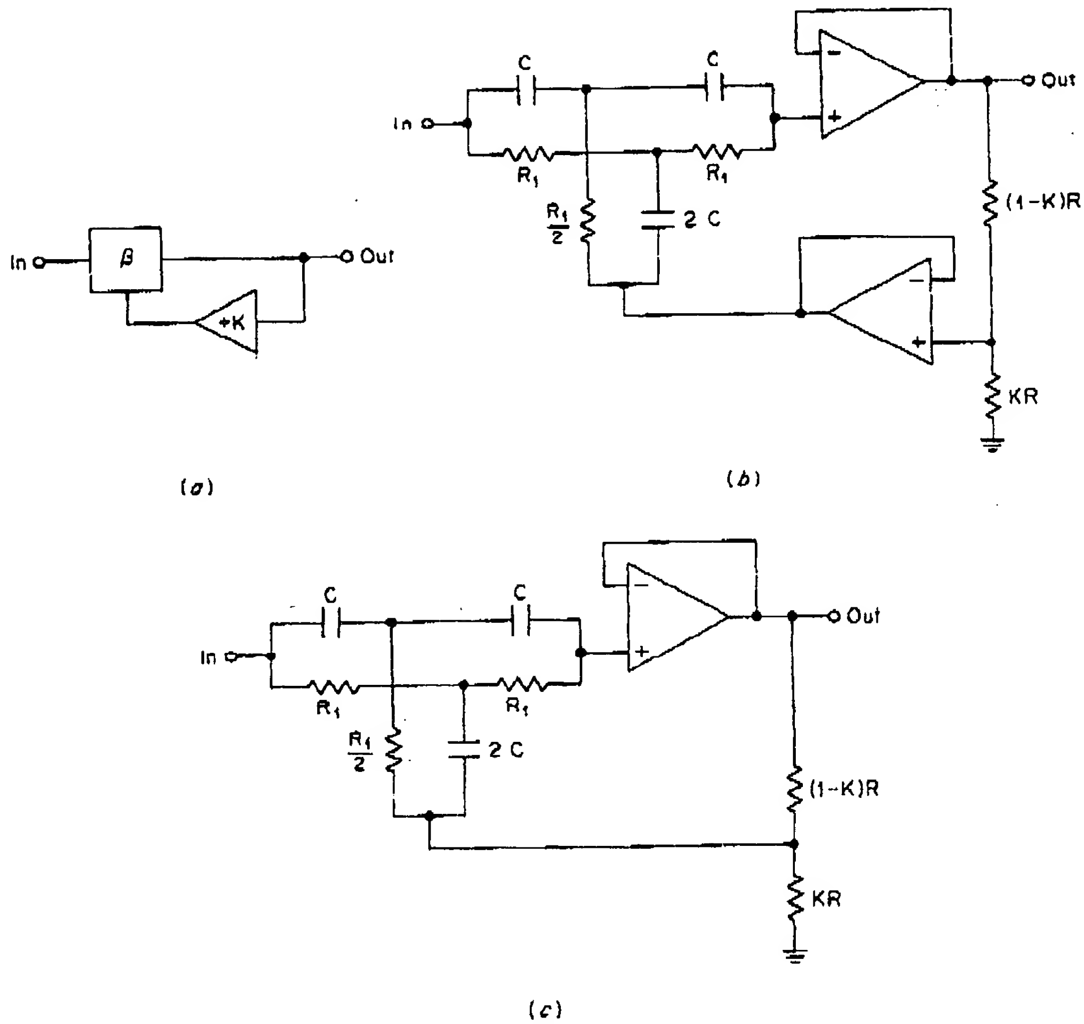


Fig. 6-27 Twin-T with positive feedback: *(a)* block diagram; *(b)* circuit realization; *(c)* simplified configuration $R_1 \gg (1-K)\,R$.

Twin-T with Positive Feedback The twin-T has gained widespread usage as a general-purpose null network. However, a major shortcoming is a fixed Q of ¼. This limitation can be overcome by introducing positive feedback.

The transfer function of the circuit of figure 6-27*a* can be derived as

$$T(s)=\frac{\beta}{1+K(\beta-1)} \tag{6-77}$$

If β is replaced by equation (6-75), the transfer function of a twin-T, the resulting circuit transfer function expression becomes

$$T(s)=\frac{s^2+\omega_0^2}{s^2+4\omega_0(1-K)s+\omega_0^2} \tag{6-78}$$

The corresponding Q is

$$Q=\frac{1}{4(1-K)} \tag{6-79}$$

By selecting a positive K of < 1 and sufficiently close to unity, th circuit Q can be dramatically increased. The required value of K can be determined by

$$K = 1 - \frac{1}{4Q} \tag{6-80}$$

The block diagram of figure 6-27*a* can be implemented using the circuit of figure 6-27*b*, where R is arbitrary. By choosing C and R so that $R_1 >> (1 - K)R$, the circuit may be simplified to the configuration of figure 6-27*c*, which uses only one amplifier.

The attenuation at any bandwidth is given by

$$A_{dB} = 10 \log \left[1 + \left(\frac{f_0}{Q \times BW_{x\ dB}} \right)^2 \right] \tag{6-81}$$

Equation (6-81) is the general expression for the attentuation of a single band-reject section where the resonant frequency and notch frequency are identical (i.e., $f_r = f_\infty$). The attenuation formula can be expressed in terms of the 3-dB bandwidth as follows:

$$A_{dB} = 10 \log \left[1 + \left(\frac{BW_{3\ dB}}{BW_{x\ dB}} \right)^2 \right] \tag{6-82}$$

The attenuation characteristics can also be determined from the frequency-response curve of a normalized $n = 1$ Butterworth low-pass filter (see figure 2-34) by using the ratio $BW_{3\ dB}/BW_{x\ dB}$ for the normalized frequency.

The twin-T in its basic form or in the positive-feedback configuration is widely used for single-section band-reject sections. However, it suffers from the fact that tuning cannot be easily accomplished. Tight component tolerances may then be required to ensure sufficient accuracy of tuning and adequate notch depth. About 40- to 60-dB rejection at the notch could be expected using 1% components.

Example 6-9

REQUIRED: Design a single null network having a center frequency of 1000 Hz and a 3-dB bandwidth of 100 Hz. Also determine the attenuation at the 30-Hz bandwidth.

RESULT: (*a*) A twin-T structure with positive feedback will be used. To design the twin-T, first choose a capacitance C of 0.01 μF. The value of R_1 is given by

$$R_1 = \frac{1}{2\pi f_0 C} = \frac{1}{2\pi \times 10^3 \times 10^{-8}} = 15.9\ \text{k}\Omega \tag{6-74}$$

(*b*) The required value of K for the feedback network is calculated from

$$K = 1 - \frac{1}{4Q} = 1 - \frac{1}{4 \times 10} = 0.975 \tag{6-80}$$

where $Q = f_0/BW_{3\ dB}$

(*c*) The single amplifier circuit of figure 6-27*c* will be used. If R is chosen at 1 kΩ, the circuit requirement for $R_1 >> (1 - K)R$ is satisfied. The resulting section is shown in figure 6-28.

(*d*) To determine the attenuation at a bandwidth of 30 Hz, calculate

$$A_{dB} = 10 \log \left[1 + \left(\frac{BW_{3\ dB}}{BW_{x\ dB}} \right)^2 \right] = 10 \log \left[1 + \left(\frac{100\ \text{Hz}}{30\ \text{Hz}} \right)^2 \right] \tag{6-82}$$
$$= 10.8\ \text{dB}$$

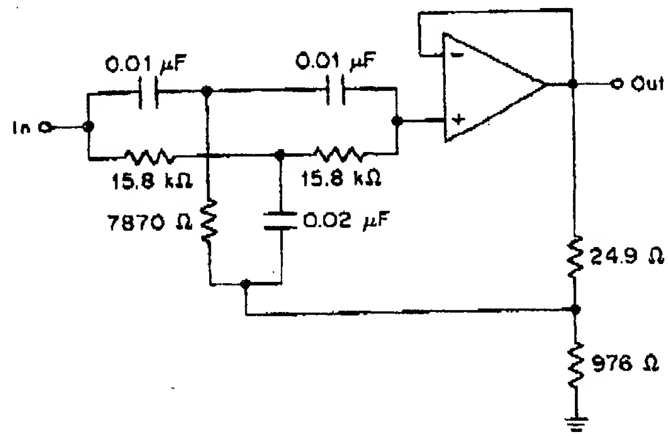


Fig. 6-28 Twin-T network of example 6-9.

Bandpass Structure Null Networks Section 6.2 under Narrow-Band Active Band-Reject Filters showed how a first-order bandpass section can be combined with a summing amplifier to obtain a band-reject circuit for transformed real poles where $f_r = f_\infty$. Three types of sections were illustrated in figure 6-23 corresponding to different Q ranges of operation. These same sections can be used as null networks. They offer more flexibility than the twin-T, since the null frequency can be adjusted to compensate for component tolerances. In addition the DABP and biquad circuits permit Q adjustment as well.

The design formulas were given by equations (6-67) through (6-73). The values of f_r and Q in the equations correspond to the section center frequency and Q, respectively.

Frequently a bandpass and band-reject output are simultaneously required. A typical application might involve separation of signals for comparison of in-band and out-of band spectral energy. The band-reject sections of figure 6-23 can each provide a bandpass output from the bandpass section along with the null output signal. An additional feature of this technique is that the bandpass and band-reject outputs will track.

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Active Filter Design Handbook

For Use with Programmable Pocket Calculators and Minicomputers

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7. BR–LQ (BAND REJECT LOW Q)

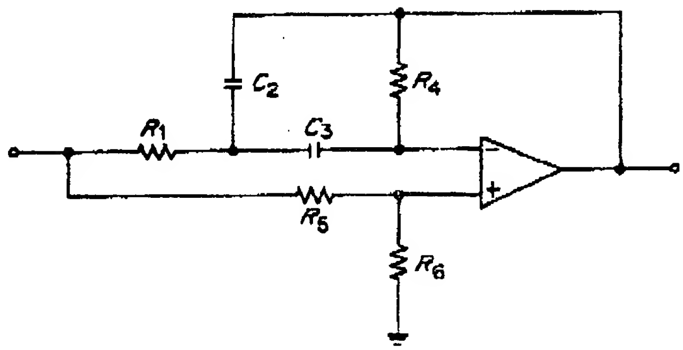

Fig. 5-7

$$T(s) = K\frac{s^2 + \omega_p^2}{s^2 + (\omega_p/q_p)s + \omega_p^2} \qquad (7a)$$

$$\omega_p^2 = \frac{1}{R_1C_2C_3R_4} \qquad (7b)$$

$$\frac{1}{R_4C_2} + \frac{1}{R_4C_3} = \frac{1}{R_1C_2}\frac{R_5}{R_6} \qquad (7c)$$

$$q_p = \frac{\sqrt{R_4/R_1}}{\sqrt{C_2/C_3} + \sqrt{C_3/C_2}} \qquad (7d)$$

$$K = \frac{R_6}{R_5 + R_6} \qquad (7e)$$

$$\mathrm{GSP} = q_p\sqrt{\frac{R_4C_3}{R_1C_2}} \qquad (7f)$$

3. BR–MQ (BAND REJECT MEDIUM Q)

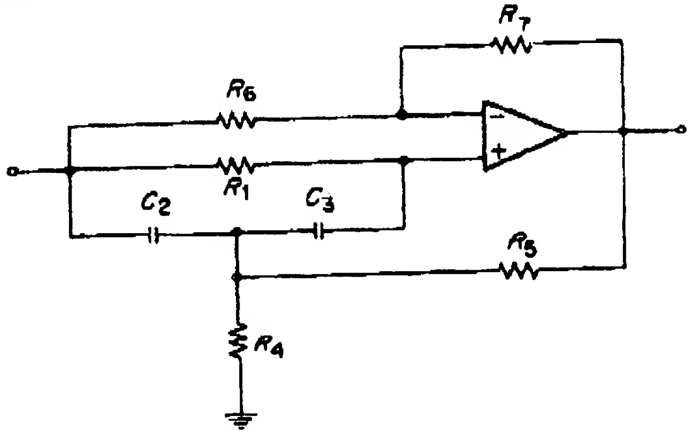

Fig. 5-13

$$T(s) = \frac{s^2 + \omega_p^2}{s^2 + (\omega_p/q_p)s + \omega_p^2} \qquad (13a)$$

$$R_p = R_4 \| R_5 \qquad (13b) \qquad\qquad \omega_p^2 = \frac{1}{R_1 C_2 C_3 R_p} \qquad (13c)$$

$$\frac{1}{R_1 C_2} + \frac{1}{R_1 C_3} = \frac{1}{R_p C_2}\frac{R_7}{R_6} \qquad (13d)$$

$$q_p = \frac{\omega_p}{1/R_1C_2 + 1/R_1C_3 + 1/R_4C_2 - (1/R_5C_2)(R_7/R_6)} \qquad (13e)$$

$$\mathrm{GSP} = q_p \frac{R_p}{R_5}\left(1 + \frac{R_7}{R_6}\right)^2 \sqrt{\frac{R_1 C_3}{R_p C_2}} \qquad (13f)$$

Tuning: (1) f_p with R_1, (2) $|T(f_p)| = 0$ with R_4, (3) q_p with R_7

20. BR–HQ (BAND REJECT HIGH Q)

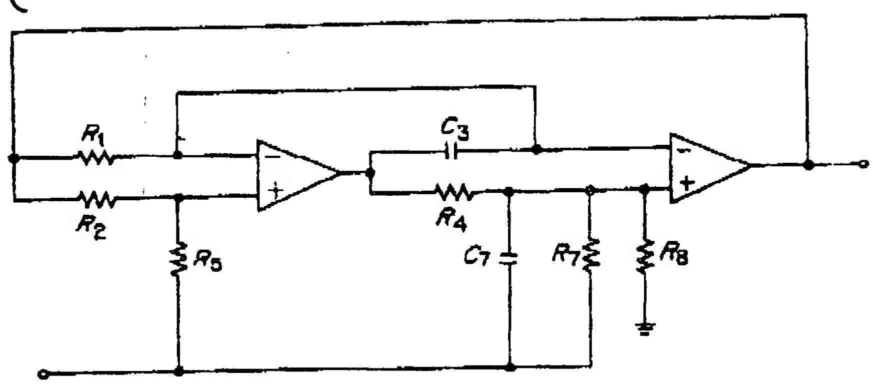

Fig. 5-20

$$T(s) = \frac{s^2 + \omega_p^2}{s^2 + (\omega_p/q_p)s + \omega_p^2} \qquad (20a)$$

$$\omega_p^2 = \frac{R_2}{R_1 R_4 R_5 C_3 C_7} \qquad (20b)$$

$$q_p = \omega_p C_7 \frac{R_7 R_8}{R_7 + R_8} \qquad (20c)$$

$$R_2 R_7 = R_5 R_8 \qquad (20d)$$

Tuning: (1) f_p with R_4, (2) $|T(f_p)| = 0$ and q_p with R_7 and R_8 (iterative)